BY JIM WILLIAMS • LINEAR TECHNOLOGY CORP

Designing instrumentation circuitry with rms/dc converters

RMS CONVERTERS RECTIFY AVERAGE RESULTS.

Using rms to measure waveforms furnishes the most accurate amplitude information (**Reference 1**). Rectify-and-average schemes, which you usually calibrate to a sine wave, are accurate for only one waveshape, however. Departures from this waveshape result in pronounced errors. Although accurate, rms conversion often entails limited bandwidth, restricted range, complexity, and difficult-to-characterize dynamic and static errors. Recent developments address these issues and also improve accuracy. **Table 1** at the Web version of this article at www.edn.com/ms4228 shows Linear Technology's (www.linear.com) LTC1966/LTC1967/LTC1968 device family. The devices feature low-frequency accuracy, including linearity and gain error, of 0.5% and 1% error at bandwidths extending to 500 kHz. These converters employ a sigma-delta-based computational scheme to achieve their performance.

Figure 1's pinout descriptions and basic circuits reveal an easily applied device. An output filter capacitor is all that is necessary to form a functional rms/dc converter. The **figure** shows split- and single-supply-powered variants. Such ease of implementation invites a broad range of application; examples begin with **Figure 2**.

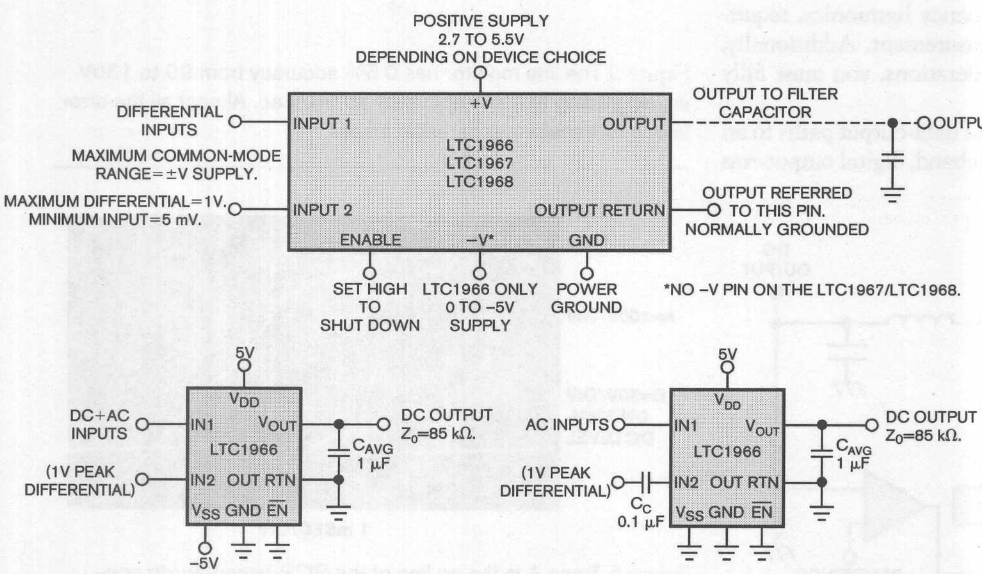


Figure 1 The pinout descriptions (top) and basic circuits (bottom) include the rms converter's pin functions and application circuits. The pins' descriptions are equivalent in all the devices, with only minor differences.

ISOLATED POWER-LINE MONITOR

Figure 2's ac-power-line monitor has 0.5% accuracy over a sensed 90 to 130V-ac input and provides a safe, fully isolated output. Conversion of rms provides accurate reporting of ac-line voltage, regardless of waveform distortion, which is common. T_1's ratio divides down the ac-line voltage. An isolated and reduced potential appears across T_1's secondary, B, at which it resistively scales and presents itself to IC_1's input. Power for IC_1 comes from T_1's secondary, A, which you rectify, filter, and zener-regulate to dc. IC_2 provides a numerically convenient output from gain. You can increase accuracy by biasing T_1 to an optimal loading point, which the relatively low-resistance-divider values facilitate. Similarly, although IC_1 and IC_2 can operate from one supply, split supplies maintain symmetrical T_1 loading. You calibrate the circuit by adjusting the 1-kΩ trim for 1.20V output with the ac line at 120V ac. You make this adjustment using a variable-ac-line transformer and a floating rms voltmeter (see **sidebar** "AC-measurement and signal-handling practice" at the Web version of this article at www.edn.com/ms4228 for recommendations on rms voltmeters and other ac-measurement-related gossip).

Figure 3's error plot shows 0.5% accuracy from 90 to 130V ac, degrading to 1.4% at 140V ac. The beneficial effect of trimming at 120V ac is evident; trimming at full-scale would result in larger overall error, primarily due to nonideal-transformer behavior. Note that the data is specific to the transformer. Substitution for T_1 necessitates circuit-value changes and recharacterization.

FULLY ISOLATED

RMS/dc converters commonly require accurate rms-amplitude measurement of an SCR's (silicon-controlled rectifier's) chopped ac-line waveforms. The

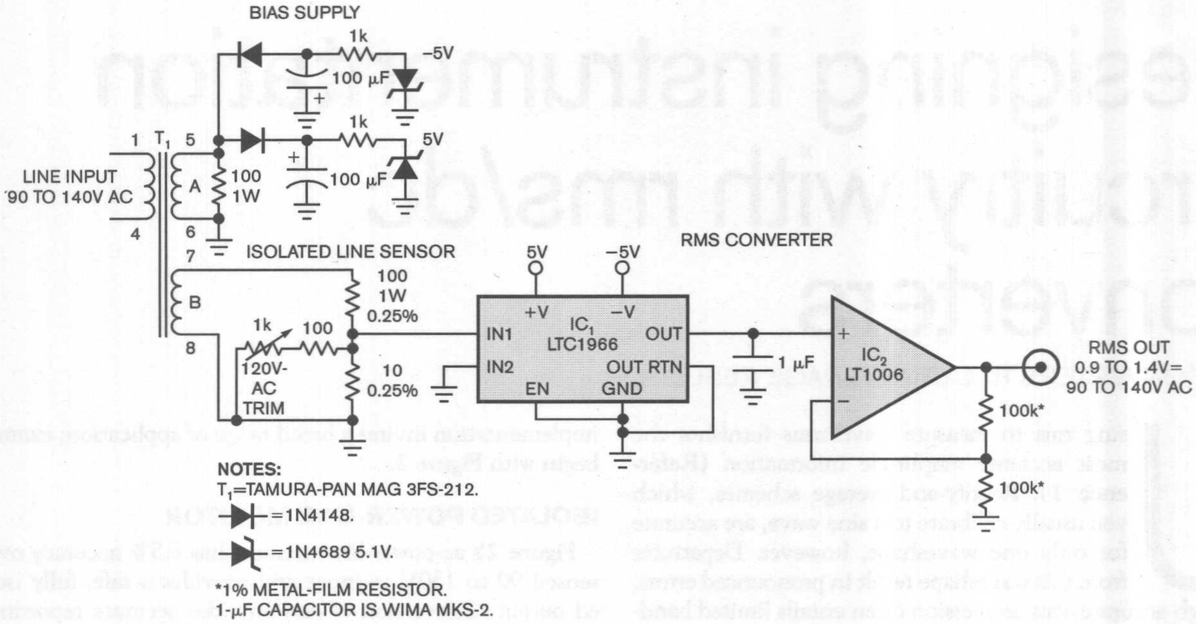


Figure 2 This isolated power-line monitor senses with a transformer and provides 0.5% accuracy from 90 to 130V-ac input. Loading the transformer secondary optimizes the voltage conversion's linearity.

SCR's fast sine-wave switching complicates this measurement because this speed introduces odd waveshapes with high-frequency harmonic content. **Figure 4**'s conceptual SCR-based ac/dc converter is typical. The SCRs alternatively chop the 220V-ac line, responding to a loop-enforced, phase-modulated trigger to maintain a dc output. **Figure 5**'s waveforms show typical operation. Trace A represents one ac-line phase, and Trace B represents the SCR cathodes. The SCR's irregularly shaped waveform contains dc and high-frequency harmonics, requiring wideband rms conversion for measurement. Additionally, for safety and system-interface considerations, you must fully isolate the measurement.

Figure 6 provides isolated power and data-output paths to an rms/dc converter, permitting safe, wideband, digital output-rms

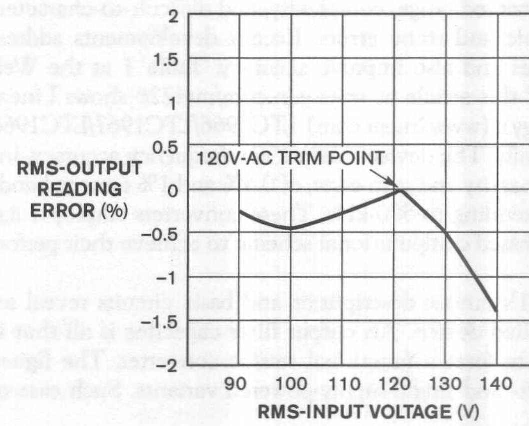


Figure 3 The line monitor has 0.5% accuracy from 90 to 130V ac, degrading to 1.4% accuracy at 140V ac. Almost all the error is due to transformer parasitic losses.

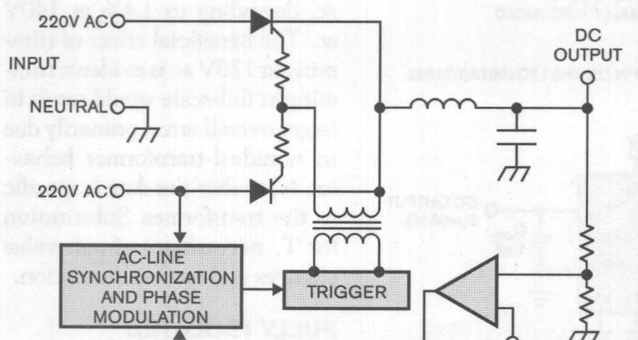


Figure 4 This ac/dc converter is typical of SCR-based designs. Feedback directs the SCR, which synchronizes with the ac line. The SCR's trigger-phase modulation controls the dc output.

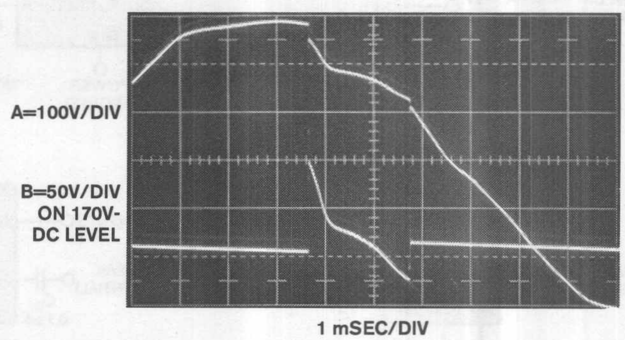


Figure 5 Trace A is the ac line of the SCR-based ac/dc converter. Trace B is the waveform at the SCR cathode. It contains dc and high-frequency harmonics that require wideband rms measurement to ensure accurate regulation.

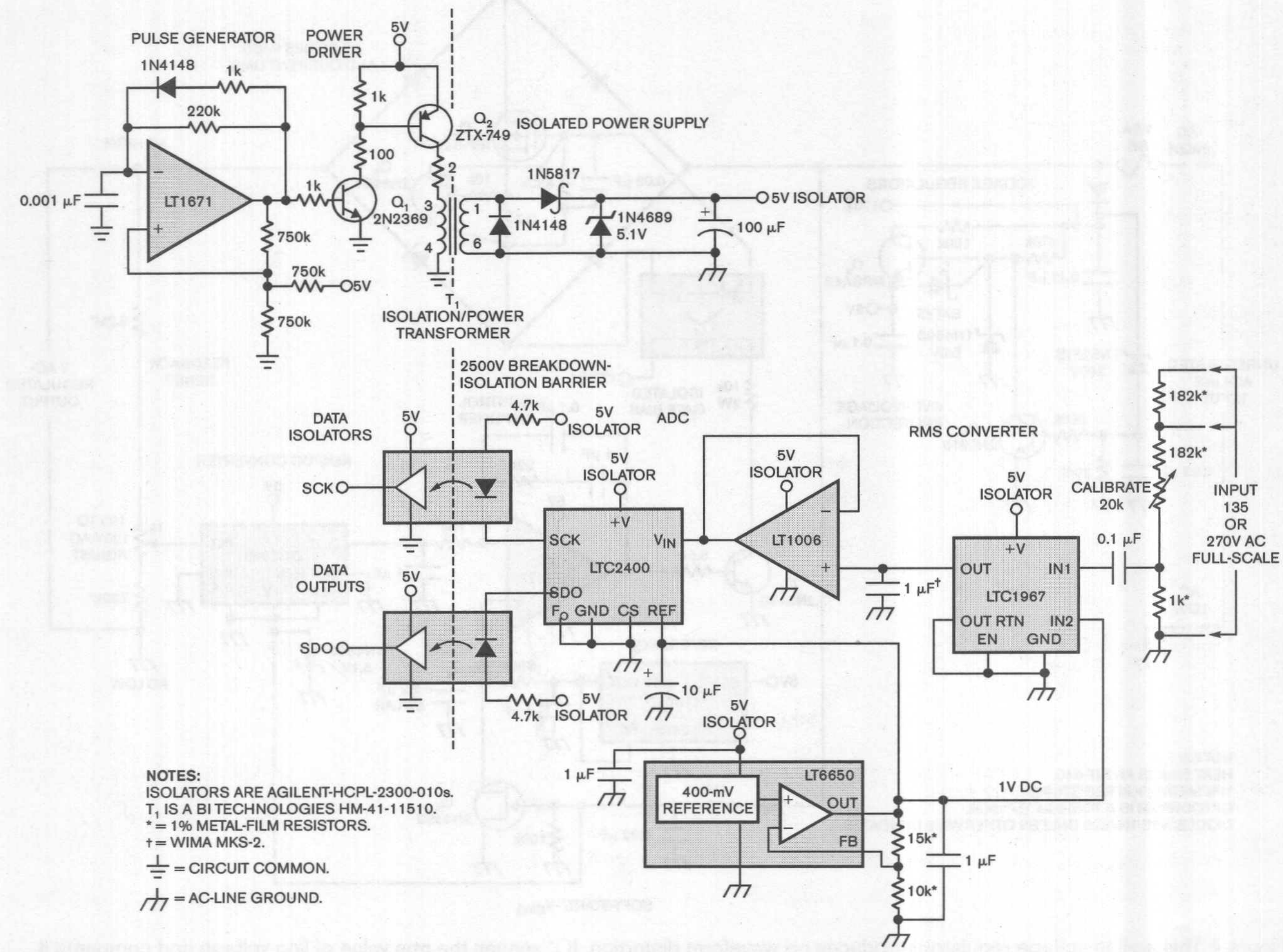


Figure 6 Optoisolators provide a safe, low-voltage digital output for this wideband-rms-measurement circuit. T_1 and associated circuitry provide isolated power for the rms converter, and a resistive divider performs high-voltage ac sensing. An ADC provides a serial output to the optoisolators. The accuracy of this circuit is 1% over a 200-kHz bandwidth.

measurement. A pulse-generator-configured comparator combines with Q_1 and Q_2 to drive T_1, resulting in isolated 5V power at T_1's rectified, filtered, and zener-regulated output. The rms/dc converter senses either 135 or 270V-ac full-scale inputs through a resistive divider. The converter's dc output feeds a self-clocked, serially interfaced ADC; optocouplers convey output data across the isolation barrier. The LTC6650 provides a 1V reference to the ADC and biases the rms/dc converter's inputs to accommodate the voltage divider's ac swing. You accomplish calibration by adjusting the 20-kΩ trim and noting that output data agrees with the input ac voltage. Circuit accuracy is within 1% in a 200-kHz bandwidth.

LOW-DISTORTION AC-LINE RMS REGULATOR

Almost all functioning ac-line-voltage regulators rely on some form of waveform chopping, clipping, or interruption. This requirement promotes efficiency but introduces waveform distortion, which is unacceptable in some applications. **Figure 7** regulates the ac line's rms value within 0.25% over wide input swings and introduces no distortion. It accomplishes this task by continuously controlling the conductivity of a series-pass MOSFET in the ac line's path. Enclosing the MOSFET in a diode bridge permits it to operate during both ac-line polarities.

You apply the ac-line voltage to the Q_2-diode bridge. A calibrated variable-voltage divider senses this bridge and feeds IC_1. You route IC_1's output, representing the regulated line's rms value, to control amplifier IC_2 and compare it with a reference. IC_2's output biases Q_1, controlling drive to a photovoltaic optoisolator. The optoisolator's output voltage provides level-shifted bias to diode-bridge-enclosed Q_2, closing a control loop, which regulates the output's rms voltage against ac-line and -load shifts. RC components in IC_2's local feedback path stabilize the control loop. The loop operates Q_2 in its linear region, much like a common low-voltage dc linear regulator. The result is the absence of introduced distortion at the expense of lost power. Heat dissipation constrains the available output power. For example, when you set the output adjustment to regulate 10V below the normal input, Q_2 dissipates about 10W at 100W output. You can improve this figure, however. The circuit regulates for $V_{IN} \geq 2V$ above V_{OUT}, but operation in this region risks regulation dropout as V_{IN} varies.

Circuit details include JFET Q_5 and associated components. The passive components associated with Q_5's gate form a slow

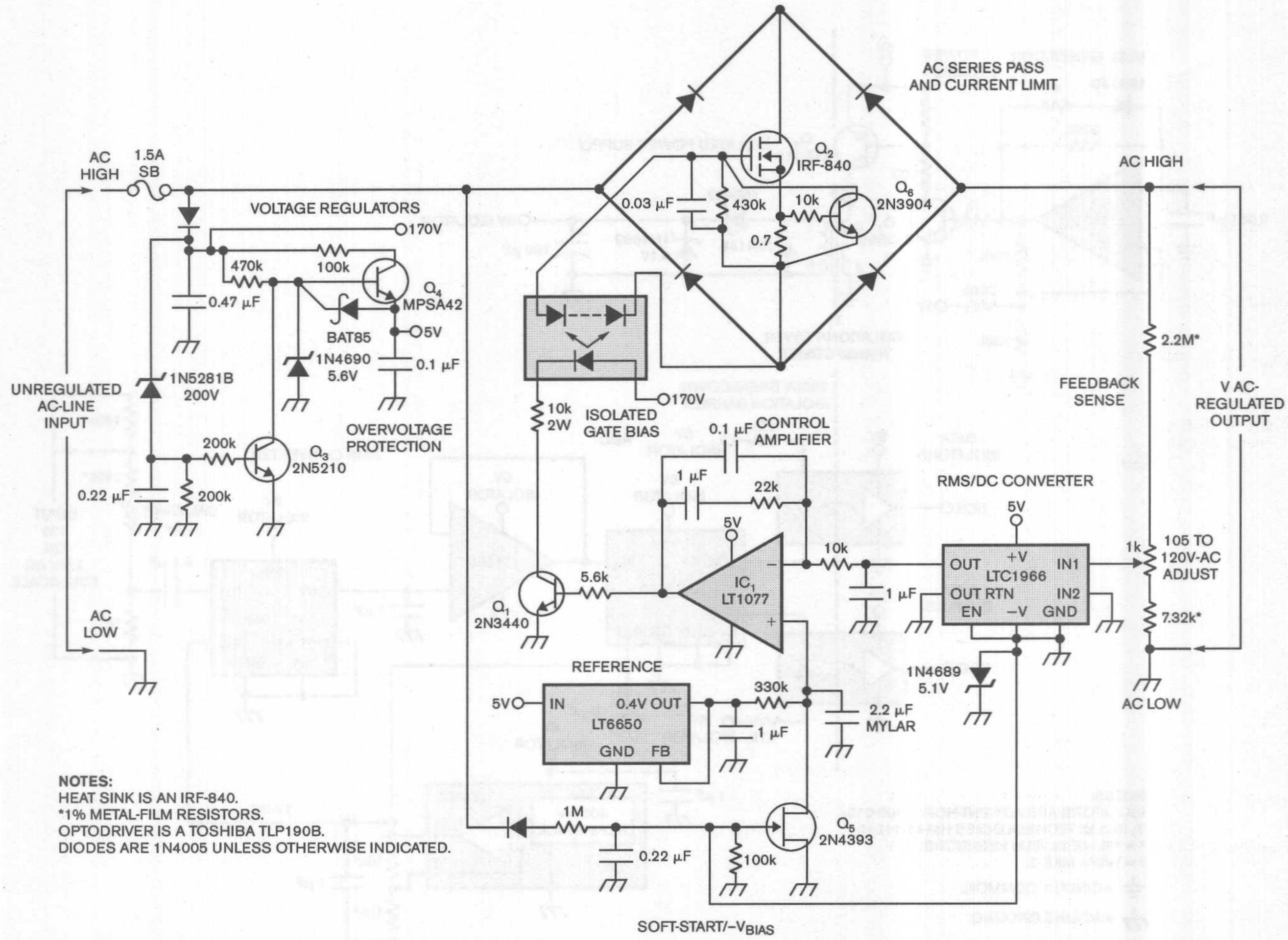


Figure 7 This ac-line-voltage regulator introduces no waveform distortion. IC_2 senses the rms value of line voltage and compares it with a reference. IC_2 then biases the photovoltaic optocoupler through Q_1. Q_2 sets the diode bridge's conductivity and closes the control loop. The input voltage must be 2V higher than the output voltage to maintain regulation.

turn-on negative supply for IC_1. They also provide gate bias for Q_5, a soft-start transistor that prevents abrupt ac power application to the output at start-up. When power is off, Q_5 conducts, holding IC_2's positive input low. When you apply power, IC_1 initially has a 0V reference, causing the control loop to set the output at zero. As the 1 MΩ, 0.22-µF combination charges, Q_5's gate moves negative, causing its channel conductivity to gradually decay. Q_5 ramps off, IC_2's positive input moves smoothly toward the LT6650's 400-mV reference, and the ac output similarly ascends toward its regulation point. Current sensor Q_6, measuring across the 0.7Ω shunt, limits output current to approximately 1A. At normal line inputs of 90 to 135V ac, Q_4 supplies 5V operating bias to the circuit. If line voltage rises beyond this point, Q_3 comes on, turning off Q_4 and shutting down the circuit.

GAIN-OF-1000 PREAMPLIFIERS

The preceding circuits furnish high-level inputs to the rms converter. Many applications lack this advantage and require some form of preamplifier. High gain preamplification for the rms converter requires more attention than you might suppose. The preamplifier must have low offset error because the rms converter (desirably) processes dc as legitimate input. More subtly, the preamplifier must have far more bandwidth than is immediately apparent. The amplifier's −3-dB bandwidth is of interest, but its closed-loop 1%-amplitude-error bandwidth must be high enough to maintain accuracy over the rms converter's 1%-error passband. This requirement is not trivial, because very high open-loop gain at the maximum frequency of interest is necessary to avoid inaccurate closed-loop gain.

Figure 8 shows a gain-of-1000 preamplifier that preserves the LTC1966's dc to 6-kHz, 1% accuracy. The amplifier may be either ac- or dc-coupled to the rms converter. The 1-mV full-scale input splits into high- and low-frequency paths. IC_1 and IC_2, which are both ac-coupled, take a cascaded, high-frequency gain of 1000. Chopper-stabilized IC_3 which is dc-coupled, also has a gain of 1000, but its RC-input filter restricts it to operate only at dc and low frequency. Assuming the switch is set to dc+ac, high- and low-frequency-path information recombine at the rms converter. The high-frequency path's 650-kHz, −3-dB response combines with the low-frequency section's microvolt-level offset to preserve the rms converter's dc to 6-kHz 1% error. If you require only ac response, set the switch to the appropriate position. The minimum processable input, which the circuit's noise floor sets, is 15 µV.

The LTC1968, with a 500-kHz, 1%-error bandwidth, poses a

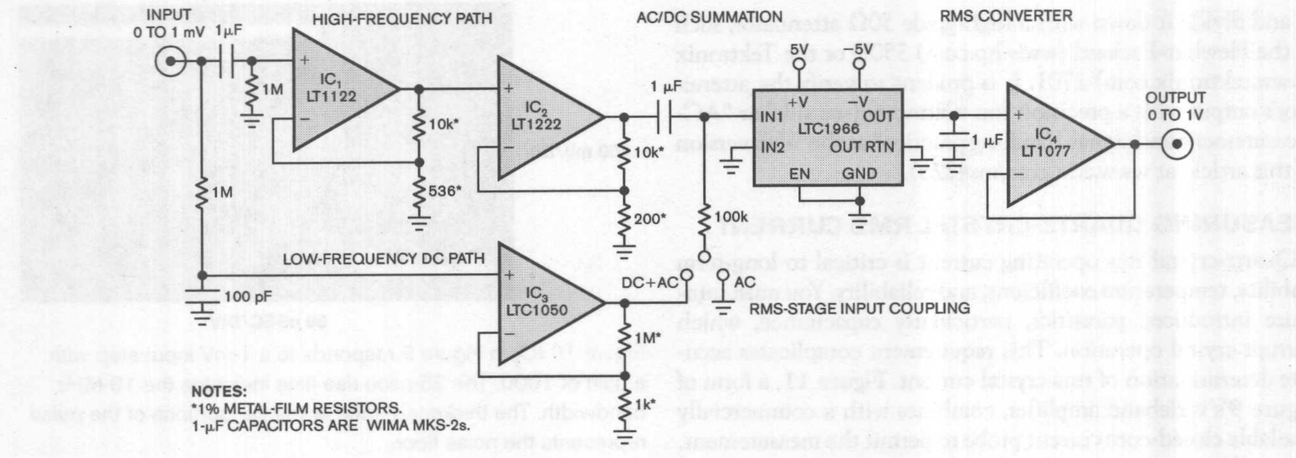
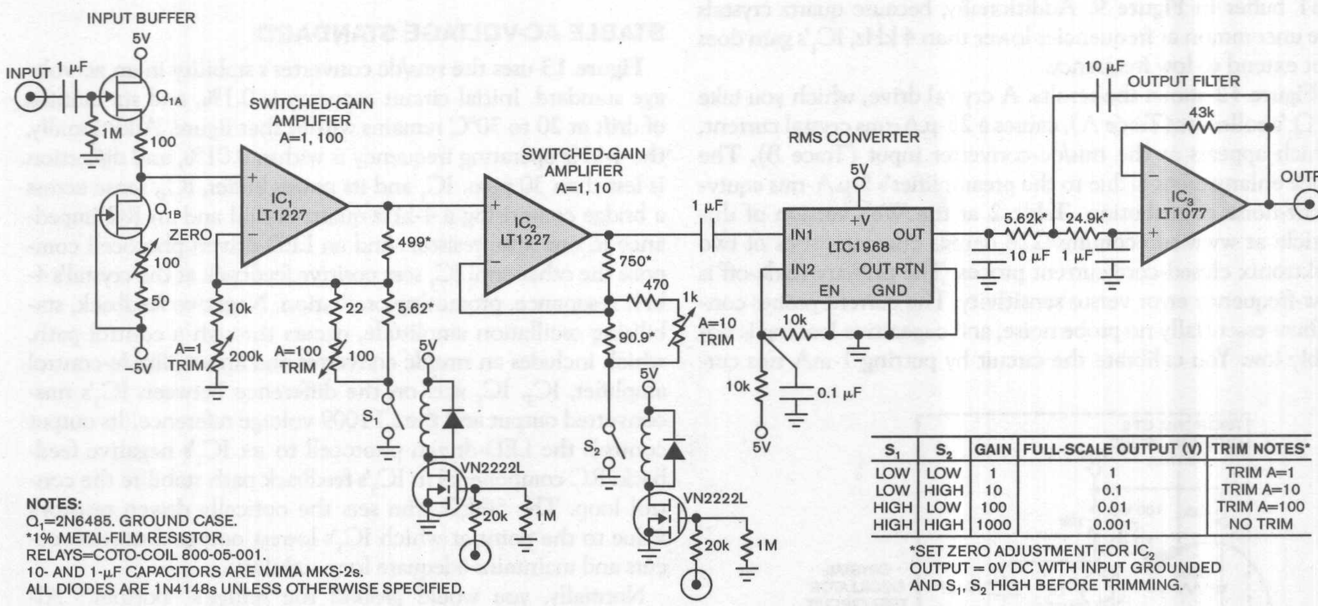


Figure 8 This gain-of-1000 preamplifier allows rms-to-dc conversion with 1-mV full-scale sensitivity. The input splits into high- and low-frequency paths that recombine at the rms converter. The amplifier's 650-kHz, −3-dB bandwidth preserves the rms converter's 6-kHz, 1%-error bandwidth. The noise floor of this circuit is 15 µV.



S_1	S_2	GAIN	FULL-SCALE OUTPUT (V)	TRIM NOTES*
LOW	LOW	1	1	TRIM A=1
LOW	HIGH	10	0.1	TRIM A=10
HIGH	LOW	100	0.01	TRIM A=100
HIGH	HIGH	1000	0.001	NO TRIM

*SET ZERO ADJUSTMENT FOR IC_2 OUTPUT = 0V DC WITH INPUT GROUNDED AND S_1, S_2 HIGH BEFORE TRIMMING.

Figure 9 This switched-gain, 10-MHz, −3-dB ac preamplifier preserves the LTC1968's 500-kHz, 1%-error bandwidth. The decade-ranged gains allow a 1-mV full-scale reading with a 20-µV noise floor. The JFET-input stage provides high input impedance. AC coupling and a third-order Sallen-Key filter maintain 1% accuracy down to 10 Hz.

significant challenge for an accurate preamplifier, but the circuit in **Figure 9** meets the requirement. This design features decade-ranged gain to 1000 with a 1%-error bandwidth beyond 500 kHz, preserving the rms converter's 1%-error bandwidth. Its 20-µV noise floor maintains wideband performance at microvolt-level inputs. Q_{1A} and Q_{1B} form a low-noise buffer, permitting high-impedance inputs. IC_1 and IC_2, which are both gain-switchable, take cascaded gain in accordance with the **figure**'s table. You set the gains using reed relays, which a 2-bit code controls. IC_2's output feeds the rms converter, and a Sallen-Key active filter smoothes the converter's output. The circuit maintains 1% error over a 10-Hz to 500-kHz bandwidth at all gains due to the preamplifier's −3-dB, 10-MHz bandwidth. You can eliminate the 10-Hz, low-frequency restriction with a dc-stabilization path similar to the one in **Figure 8**, but you would have to switch its gain in concert with the IC_1-IC_2 path.

Figure 10 shows preamplifier response to a 1-mV input step at a gain of 1000. IC_2's output is singularly clean, with trace thickening in the pulse's flat portions due to the 20-µV noise floor. The 35-nsec rise time indicates a 10-MHz bandwidth. To calibrate this circuit, first set S_1 and S_2 high, ground the input, and trim the zero adjustment for 0V dc at IC_2's output. Next, set S_1 and S_2 low, apply a 1V, 100-kHz input, and trim A=1 for unity gain, which you measure at the circuit output, in accordance with the table in **Figure 9**. Continue this procedure for the remaining three gains in the table. A good way of generating the required accurate low-level inputs is to set a 1V-ac lev-

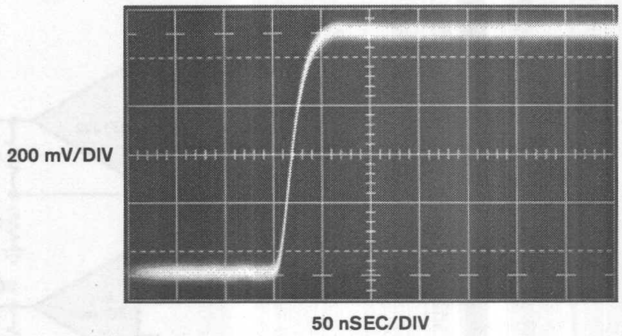


Figure 10 IC_2 in Figure 9 responds to a 1-mV input step with a gain of 1000. The 35-nsec rise time indicates the 10-MHz bandwidth. The thickened trace at the flat portions of the pulse represents the noise floor.

el and divide it down with a high-grade 50Ω attenuator, such as the Hewlett-Packard (www.hp.com) 350D or the Tektronix (www.tektronix.com) 2701. It is prudent to verify the attenuator's output with a precision rms voltmeter (see **sidebar** "AC-measurement and signal-handling practice" at the Web version of this article at www.edn.com/ms4228).

MEASURING QUARTZ-CRYSTAL RMS CURRENT

Quartz-crystal rms operating current is critical to long-term stability, temperature coefficient, and reliability. You must minimize introduced parasitics, particularly capacitance, which corrupt crystal operation. This requirement complicates accurate determination of rms-crystal current. **Figure 11**, a form of **Figure 9**'s wideband amplifier, combines with a commercially available closed-core current probe to permit the measurement. An rms/dc converter supplies the rms value. The quartz-crystal test circuit in dashed lines exemplifies a typical measurement situation. The Tektronix CT-2 current probe monitors crystal current and introduces minimal parasitic loading. The probe's 50Ω termination allows direct connection to IC_1 without the FET buffer in **Figure 9**. Additionally, because quartz crystals are uncommon at frequencies lower than 4 kHz, IC_1's gain does not extend to low frequency.

Figure 12 shows the results. A crystal drive, which you take at Q_1's collector (Trace A), causes a 25-μA-rms crystal current, which appears at the rms/dc-converter input (Trace B). The trace enlargement is due to the preamplifier's 5-μA-rms equivalent-noise contribution. **Table 2** at the Web version of this article at www.edn.com/ms4228 details characteristics of two Tektronix closed-core current probes. The primary trade-off is low-frequency error versus sensitivity. The current probes contribute essentially no probe noise, and capacitive loading is notably low. You calibrate the circuit by putting 1-mA rms current through the probe and adjusting the indicated trim for a 1V circuit output. To generate the 1 mA, drive a 1-kΩ, 0.1% resistor with 1V rms.

STABLE AC-VOLTAGE STANDARD

Figure 13 uses the rms/dc converter's stability in an ac-voltage standard. Initial circuit accuracy is 0.1%, and six months of drift at 20 to 30°C remains within that figure. Additionally, the 4-kHz operating frequency is within 0.01%, and distortion is less than 30 ppm. IC_1 and its power buffer, IC_3, sense across a bridge comprising a 4-kHz quartz crystal and an RC impedance in one arm; resistors and an LED-driven photocell comprise the other arm. IC_1 sees positive feedback at the crystal's 4-kHz resonance, promoting oscillation. Negative feedback, stabilizing oscillation amplitude, occurs through a control path, which includes an rms/dc converter and an amplitude-control amplifier, IC_5. IC_5 acts on the difference between IC_3's rms-converted output and the LT1009 voltage reference. Its output controls the LED-driven photocell to set IC_1's negative feedback. RC components in IC_5's feedback path stabilize the control loop. The 50-kΩ trim sets the optically driven resistor's value to the point at which IC_3's lowest output distortion occurs and maintains adequate loop stability.

Normally, you would ground the bridge's "bottom." Although this connection works, it subjects IC_1 to common-mode swings, increasing distortion due to IC_1's finite common-mode rejection versus frequency. IC_2 eliminates this concern

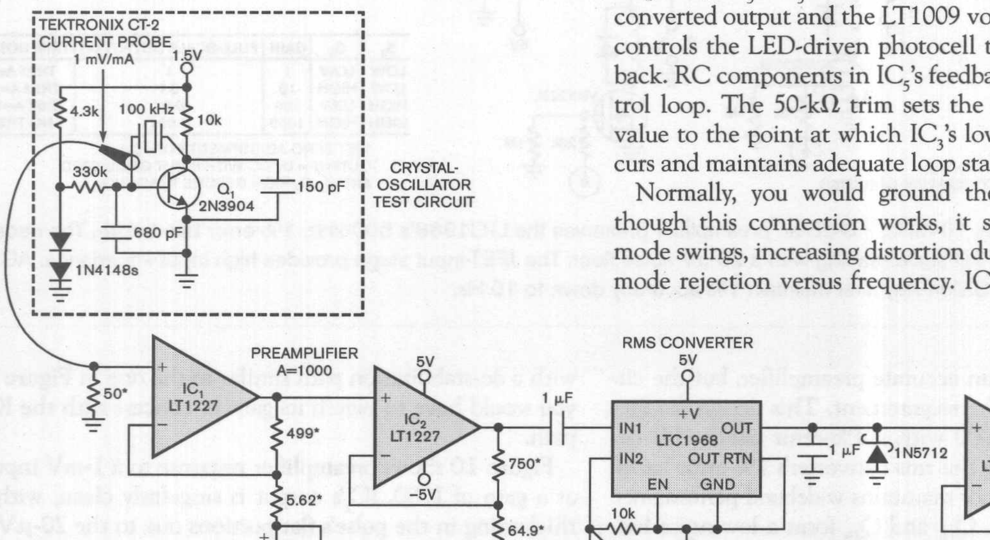


Figure 11 The circuit of Figure 9 adapts to the isolated true-rms measurement of the current in a quartz crystal. The current probe's 50Ω impedance allows the elimination of the FET-input buffer and direct connection to IC_1. The current probe does not appreciably load the crystal in this oscillator test circuit.

by forcing the bridge's midpoints and, hence, common-mode voltage to 0V but not influencing desired circuit operation. It accomplishes this task by driving the bridge bottom to force its input differential to zero. IC_2's output swing is 180° out of phase with IC_3's circuit output. This action eliminates common-mode swing at IC_1, reducing circuit output distortion by more than an order of magnitude. **Figure 14** shows the circuit's 1.414V-rms (2V peak) output in Trace A, and Trace B's distortion constituents include noise, fundamental-related residue, and second-harmonic components.

The 4-kHz crystal is a relatively large structure with a high Q factor. Normally, it would require more than 30 sec to start and arrive at full, regulated amplitude. You avoid this drawback by including the Q1-LTC201-switch circuitry. At start-up, IC_5's output goes high, biasing Q_1. Q_1's collector goes low, turning on the LTC201. This action sets IC_1's gain abnormally high, increasing bridge drive and accelerating crystal start-up. When the bridge arrives at its operating point, IC_5's output drops to a lower value, Q_1 and the LTC201 switch off, and the circuit moves into normal operation. Start-up time is several seconds.

The circuit requires trimming for amplitude accuracy and lowest distortion. You perform the distortion trim first. Adjust the trim for minimal output distortion, which you measure on a distortion analyzer. Note that the absolute lowest level of distortion coincides with the point at which control-loop gain is just adequate to maintain oscillation. As such, find this point and retreat from it into the control loop's active region. This retreat necessitates giving up about 5-ppm distortion, but you can achieve 30 ppm with good control-loop stability. You trim output amplitude with the indicated adjustment for exactly 1.414V rms (2V peak) at the circuit output.

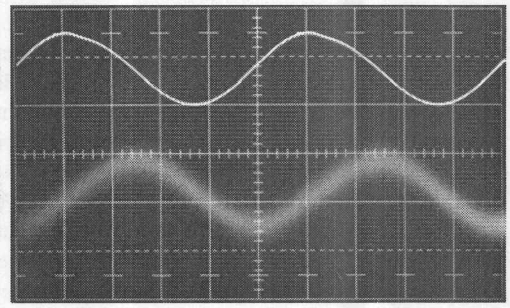


Figure 12 Trace A shows the crystal voltage, and Trace B shows the crystal current for the circuit in Figure 11. The 25-μA rms-crystal-current measurement includes the 5-μA noise-floor contribution of the preamplifier.

RANDOM-NOISE GENERATOR

Figure 15 uses the rms/dc converter in a leveled-output-random-noise generator. Noise diode D_1 ac-biases IC_1, operating at a gain of two. IC_1's output feeds a 1- to 500-kHz, switch-selectable lowpass filter. The filter output-biases the variable-gain amplifier, IC_2-IC_3. IC_2, a current-controlled transconductance amplifier, and IC_3, an output amplifier, reside on one chip. This stage takes ac gain, biases the LTC1968 rms/dc converter, and acts as the circuit's output. The rms-converter output at IC_4

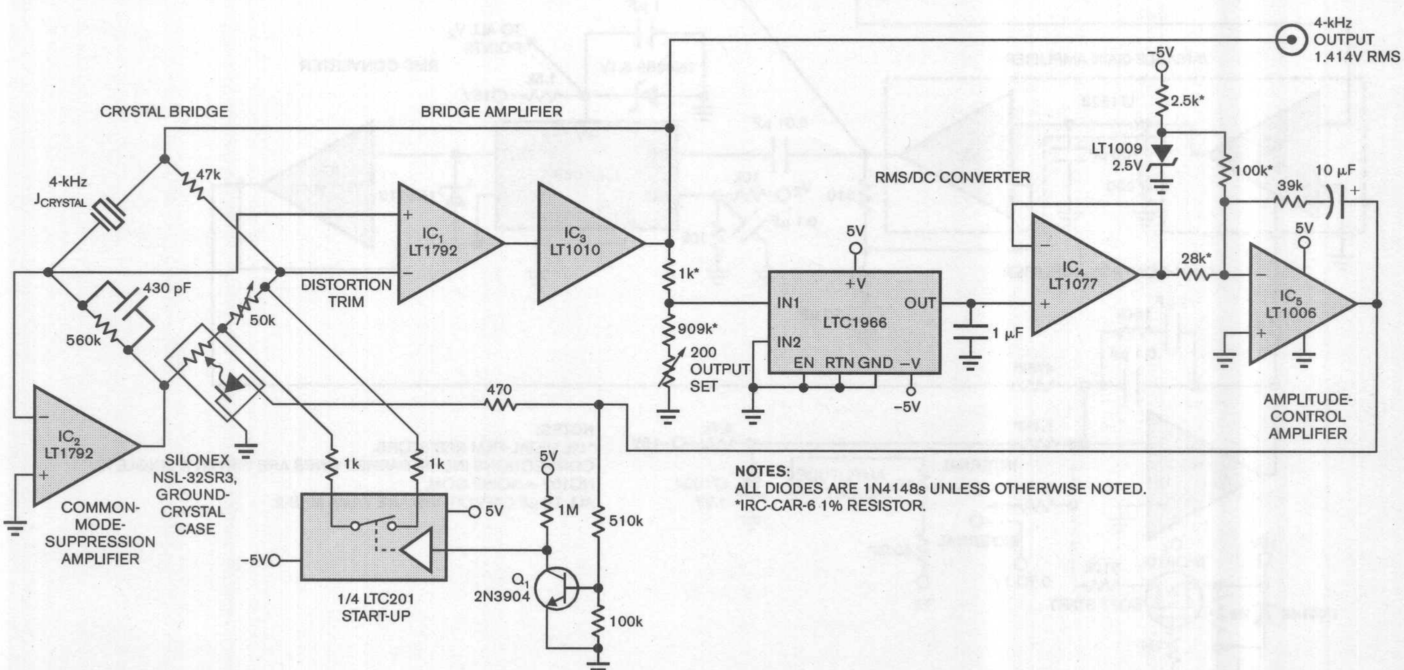


Figure 13 This quartz-stabilized sine-wave-output-ac reference has 0.1% long-term amplitude stability. The frequency accuracy is 0.01% with less-than-30-ppm distortion. The positive feedback around IC_1 causes oscillation at the crystal's resonant frequency. Amplifier IC_5 acts on the rms-amplitude output of IC_4 to supply a negative feedback to IC_1 through the bridge network that stabilizes the rms-output amplitude. The optocoupler minimizes feedback-induced distortion. Switch Q_1 closes during start-up, which ensures the rapid build up of oscillations.

feeds back to gain-control amplifier IC_5, which compares the rms value with a variable portion of the 5.1V zener potential. IC_5's output sets IC_2's gain through the 3-kΩ resistor, completing a control loop to stabilize noise-rms-output amplitude. The RC components in IC_5's local feedback path stabilize this loop. You can vary the output amplitude using the 10-kΩ potentiometer; a switch permits external voltage control. Q_1 and associated components, a soft-start circuit, prevent output overshoot at power turn-on. **Figure 16** shows circuit-output noise in the 10-kHz filter position; **Figure 17**'s spectral plot reveals essentially flat rms-noise amplitude over a 500-kHz bandwidth.

RMS-AMPLITUDE-STABILIZED LEVEL CONTROLLER

Figure 18 borrows the previous circuit's gain-control loop to stabilize the rms amplitude of an arbitrary input waveform. You apply the unregulated input to variable-gain amplifier IC_1-IC_2 which feeds IC_3. DC coupling at IC_1-IC_2 permits passage of low-frequency inputs. An rms/dc converter, comprising IC_4 and IC_6, takes IC_3's output, which feeds IC_5's gain-control amplifier. IC_5 compares the rms value with a variable reference and biases IC_1,

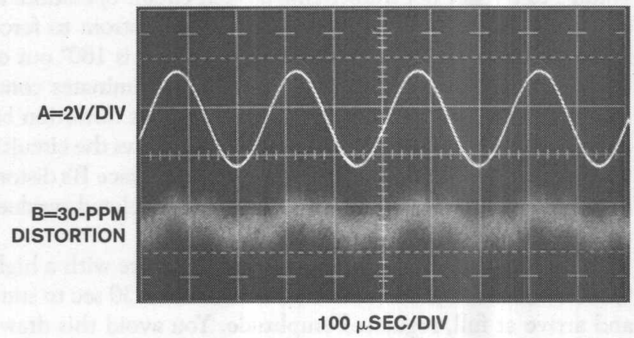


Figure 14 Trace A shows the 1.414V-rms (2V peak) reference output from IC_3. Trace B shows the 30-ppm distortion in the output. The distortion's constituents include noise, fundamental-related residue, and second-harmonic components.

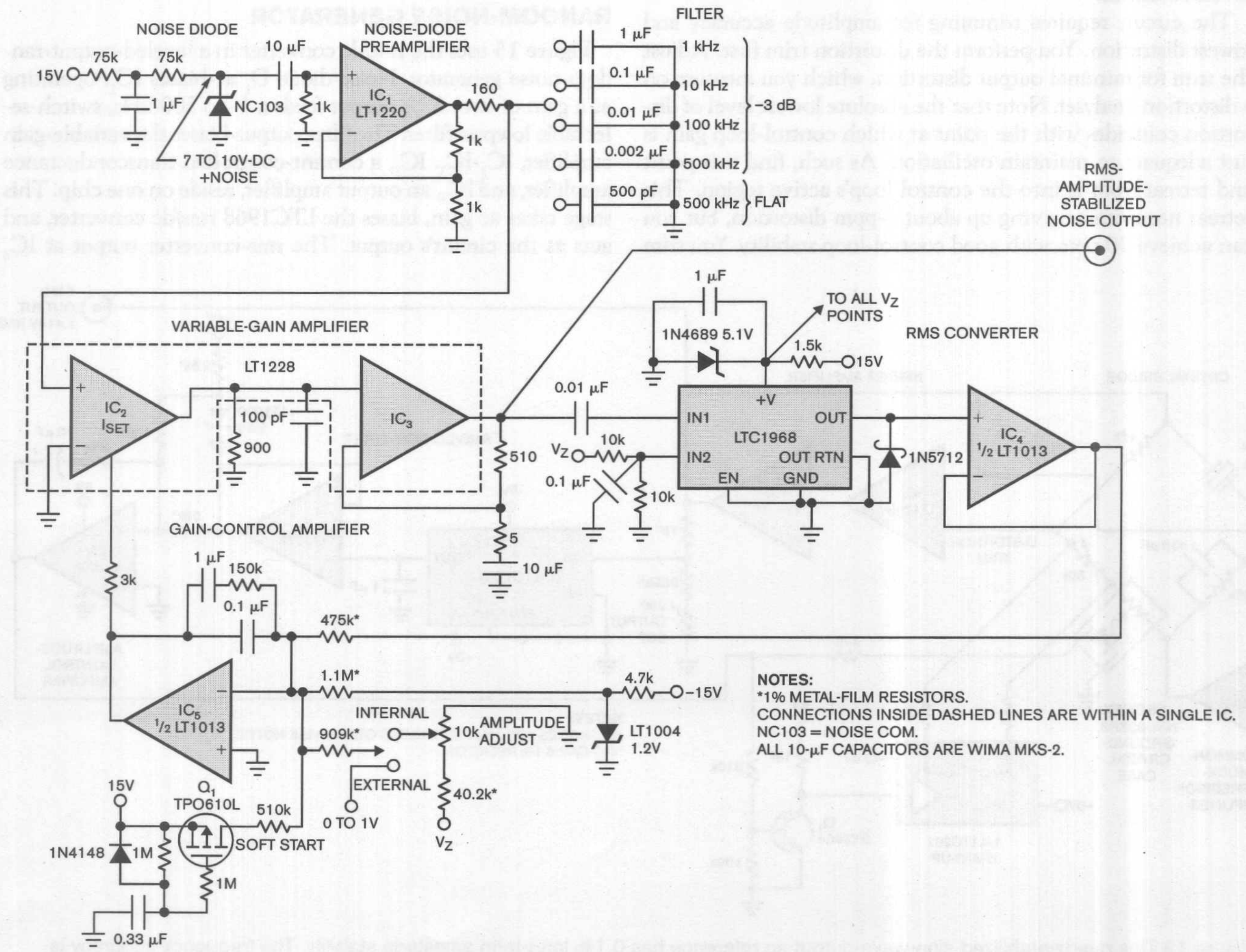


Figure 15 This circuit creates a random-noise generator with rms-leveled output. IC_1 filters and amplifies zener-diode noise. The output of the variable-gain amplifier converts to rms. The rms output feeds back to gain-control amplifier IC_5, which closes the loop to the variable-gain amplifier. A potentiometer or external input to IC_5 allows you to set the noise output to different values.

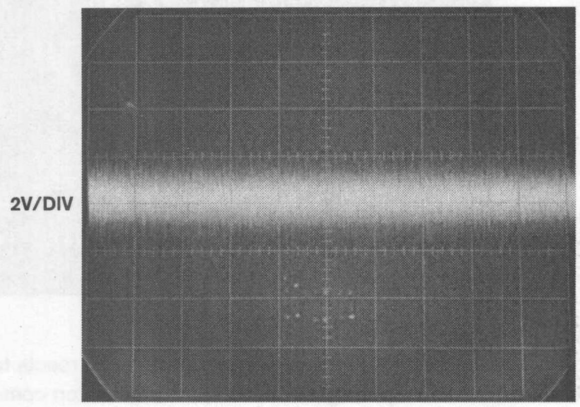


Figure 16 The output of the circuit in Figure 14 is in the 10-kHz filter position.

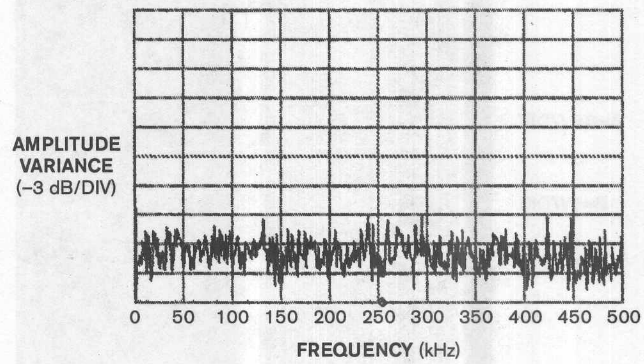


Figure 17 The amplitude over frequency for the random-noise generator is essentially flat to 500 kHz. The NC103 noise diode contributes to an even noise-spectrum distribution, and the rms converter and control loop stabilize the amplitude. The measurement sweep time is 2.8 minutes, and the resolution bandwidth is 100 Hz.

closing a gain-control loop. The 0.15-μF feedback capacitor stabilizes this loop, even for waveforms lower than 100 Hz. This feedback action maintains waveshape and stabilizes output-rms amplitude despite large variations in input amplitude. You can set the desired output level with the indicated potentiometer, or you can switch in an external control voltage.

Figure 19 shows output response (Trace B) to abrupt reference-level-setpoint changes (Trace A). The output settles within 60 msec for ascending and descending transitions. You can achieve faster response by decreasing IC_5's compensation capacitor, but the circuit would then be unable to process low-frequency waveforms. Similar considerations apply to **Figure**

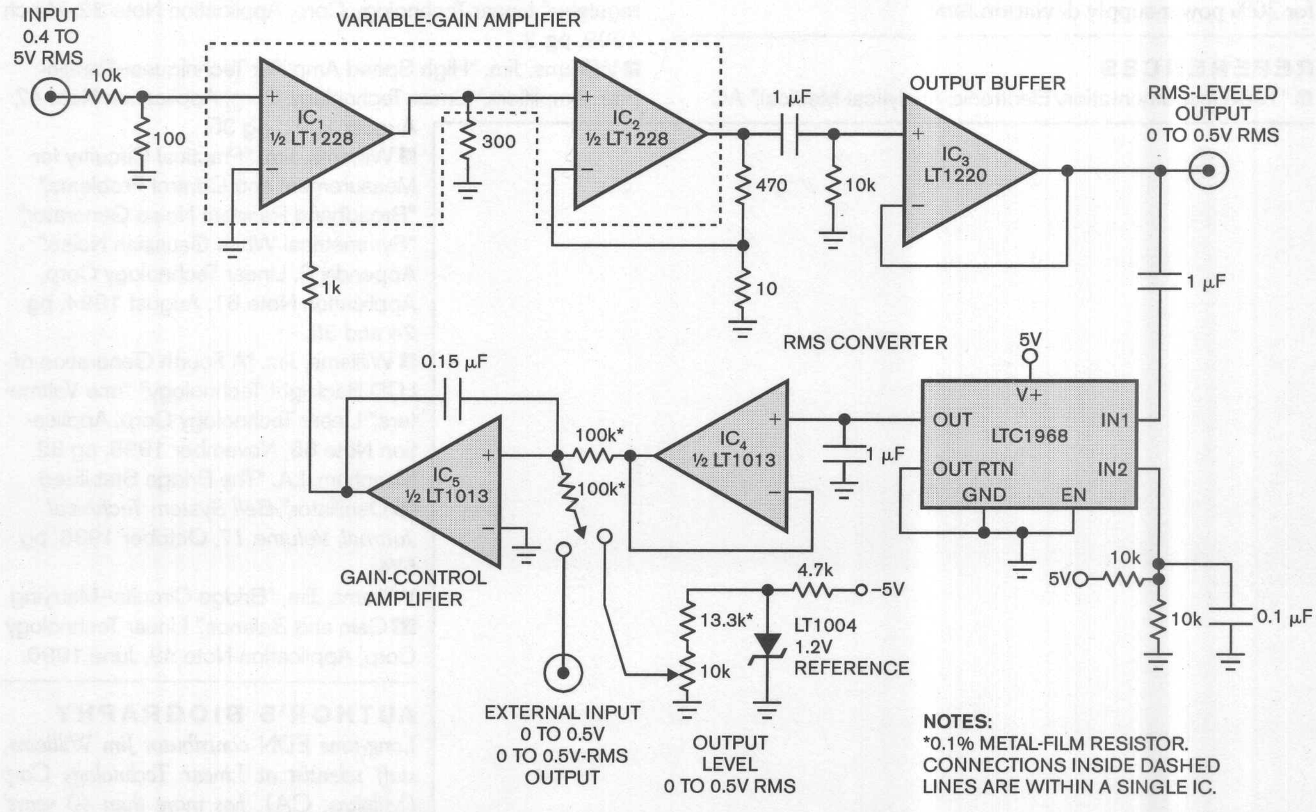


Figure 18 This rms-amplitude level-control circuit uses the gain-control loop of Figure 15. The amplifiers IC_1, IC_2, and IC_3 provide a variable-gain capability to the input section. The rms converter, IC_6, feeds back to the gain-control amplifier, IC_5, which closes the amplitude-stabilization loop. The variable-reference voltage permits a settable calibrated rms output that is amplitude-independent of the input waveshape.

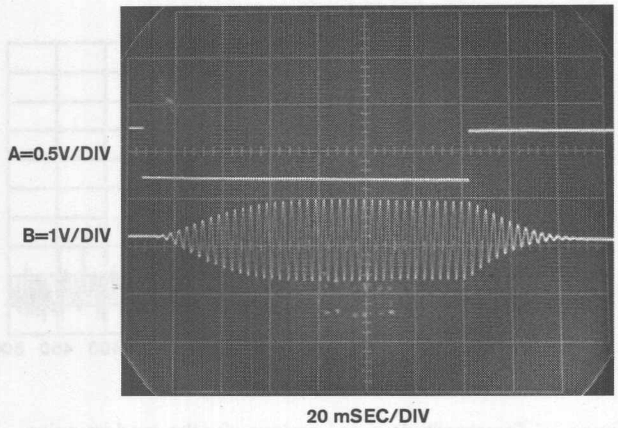


Figure 19 An abrupt change in the reference (Trace A) causes an amplitude-level-control response (Trace B). IC_5's compensation capacitor sets the settling time. This capacitor must be large enough to stabilize the loop at the lowest expected signal-input frequency.

20's response to an input-waveform step change. Trace A is the circuit's input, and Trace B is its output. The output settles in 60 msec due to IC_5's compensation. Reducing compensation value speeds response at the expense of low-frequency-waveform processing capability. Specifications include 0.1% output-amplitude stability for inputs of 0.4 to 5V rms, 1% set-point accuracy, 0.1- to 500-kHz passband, and 0.1% stability for 20% power-supply deviation. **EDN**

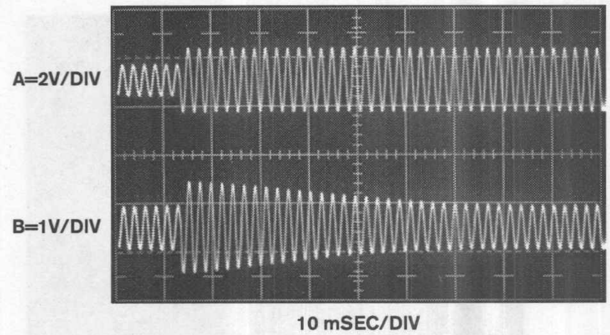


Figure 20 The amplitude-level-control output (Trace B) reacts to a step change in the input signal (Trace A). The slow loop compensation allows the overshoot, but the output settles cleanly.

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AUTHOR'S BIOGRAPHY

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AC-MEASUREMENT AND SIGNAL-HANDLING PRACTICE

Accurate ac measurement requires trustworthy instrumentation, proper signal-routing technique, parasitic minimization, attention to layout, and care in component selection. The main text describes a circuit's dc to 500-kHz, 1%-error bandwidth. These figures seem benign, but unpleasant surprises await the unwary.

Performing serious ac work requires the use of an accurate rms voltmeter. Table A lists types that Linear Technology uses in its laboratory. These thermally based, high-grade, specialized instruments make precise rms measurement. The first three entries–easy-to-use, general-purpose instruments with many ranges and features–meet almost all ac-measurement needs. The last entry is more of a component than an instrument. The A55 series of "thermal converters" provide millivolts-level outputs for various inputs. Typical input ranges are 0.5, 1, 2, and 5V rms, and each converter receives individual calibration data. They are somewhat cumbersome to use and fragile but are highly accurate. Their primary uses are as reference standards to check other instruments' performance.

AC-signal handling for high accuracy is a broad topic, involving a considerable degree of depth. Layout is critical. The most prevalent parasitic in ac measurement is stray capacitance. Keep signal-path connections short and small in area. A few picofarads of coupling into a high-impedance node can upset a 500-kHz, 1%-accuracy signal path. To the extent possible, keep impedances low to minimize parasitic-capacitive effects. Consider individual component parasitics and plan to accommodate them. Examine effects of component placement and orientation on the pc board. If a ground plane is in use, you may need to relieve it in the vicinity of critical circuit nodes or even individual components.

Keep in mind that passive components have parasitics. For example, resistors suffer shunt capacitance whose effects vary with frequency and resistor value. It is worth noting that different brands of resistors, although nominally similar, may exhibit markedly different parasitic behavior. Use capacitors in the signal path so that their outer foil connects to the less sensitive node, affording some relief from pickup and stray-capacitance-induced effects. Some capacitors have markings that indicate the outer foil terminal; others require consulting the data sheet or vendor. Avoid placing ceramic capacitors in the signal path. Their piezoelectric responses make them unsuitable for precision ac circuitry. In general, examine any component in the signal path for its potential parasitic contribution.

Treat active components, such as amplifiers, as potential error sources. In particular, ensure that there is enough open-loop gain at the frequency of interest to ensure the necessary closed-loop-gain accuracy. Margins of 100-to-1 are not unreasonable. Keep feedback values as low as possible to minimize parasitic effects.

Coaxially route signals to and from the pc board at low impedance–preferably 50Ω–for best results. In 50Ω systems, terminators and attenuators have tolerances that can corrupt a 1%-amplitude-accuracy measurement. Verify such terminator and attenuator tolerances by measurement and account for them when interpreting measurement results. Similarly, verify the accuracy of any associated instrument's 50Ω input or output impedance and account for deviations.

All this work seems painful but is an essential part of achieving 1%-accurate, 500-kHz signal integrity. Failure to observe these precautions risks degrading the rms/dc converter's system-level performance.

TABLE A PRECISION WIDEBAND-RMS VOLTMETERS

Model	Manufacturer	1V range (%)	Input	Bandwidth (MHz)	Comments
3400A/ 3400B	Hewlett-Packard	1	ac	10/20	Metered instrument, most common rms voltmeter
3403C	Hewlett-Packard	0.2	ac, ac+dc	100	Digital display, 1-μV sensitivity, 2-MHz bandwidth, decibel ranges, relative decibels
8920/ 8921A	Fluke	0.7	ac, ac+dc	20	Digital display, 10-μV sensitivity, 2-MHz bandwidth, decibel ranges, relative decibels
A55	Fluke	0.05	ac+dc	50	Set of individually calibrated thermal converters, reference standards, not for general-purpose measurement

TABLE 1 RMS/DC-CONVERTER FAMILY

Part	Typical/maximum linearity error (%)	Typical/maximum conversion gain error (%)	1%-error bandwidth (kHz)	3-dB error bandwidth	Minimum/maximum supply voltage (V)	Maximum supply current
LTC1966	0.02/0.15	0.1/0.3	6	800 kHz	2.7/±5	170 μA
LTC1967	0.02/0.15	0.1/0.3	200	4 MHz	4.5/5.5	390 μA
LTC1968	0.02/0.15	0.1/0.3	500	15 MHz	4.5/5.5	2.3 mA

TABLE 2 TEKTRONIX CURRENT PROBES

Parameter	CT-1	CT-2
Sensitivity (mV/mA)	5	1
Accuracy (%)	3	3
Low-frequency additional 1%-error bandwidth (kHz)*	98	6.4
−3-dB bandwidth	25 kHz to 1 GHz	1.2 kHz to 200 MHz
Noise floor with amplifier (μA rms)*	1	5
Capacitive loading (pF)	1.5	1.8
Insertion impedance at 10 MHz (Ω)	1	0.1

*As measured; not vendor-specified.